TRANSISTOR POWER SUPPLY FOR A PORTABLE RADAR

By Samuel A. Schwartz

I. INTRODUCTION

The transistorized power converter-inverter* described in this paper has a unique circuit that uses a power oscillator and a power amplifier. It provides 70 W of power for all the circuitry of a small portable radar set, including microwave and cathode-ray-tube heaters, motor power and multiple d-c supplies. The radar set is a small X-band navigational radar constructed at Fairchild Semiconductor and described in other application notes.**

*The term "converter" normally refers to power conversion from d-c to d-c whereas "inverter" refers to conversion from d-c to a-c. A combination converter-inverter accommodates both a-c and d-c outputs from one unit.

**APP-76 through APP-81.

The radar is powered from a 28 V d-c source; however, the design is versatile and can be adapted to either 12 V d-c or 115 V a-c.

Switching-type power converters and inverters generate high transient currents that could interfere with the proper operation of ancillary equipment. The problem has been circumvented in this converter by synchronizing its switching rate with the sweep generator of the radar so that the sweep interval is completed before the power switching takes place and hence the effect of transient currents is never observed.

The design, which is conservative, has four Fairchild 2N2892 30-W planar silicon power transistors in the power switching section. Two are used

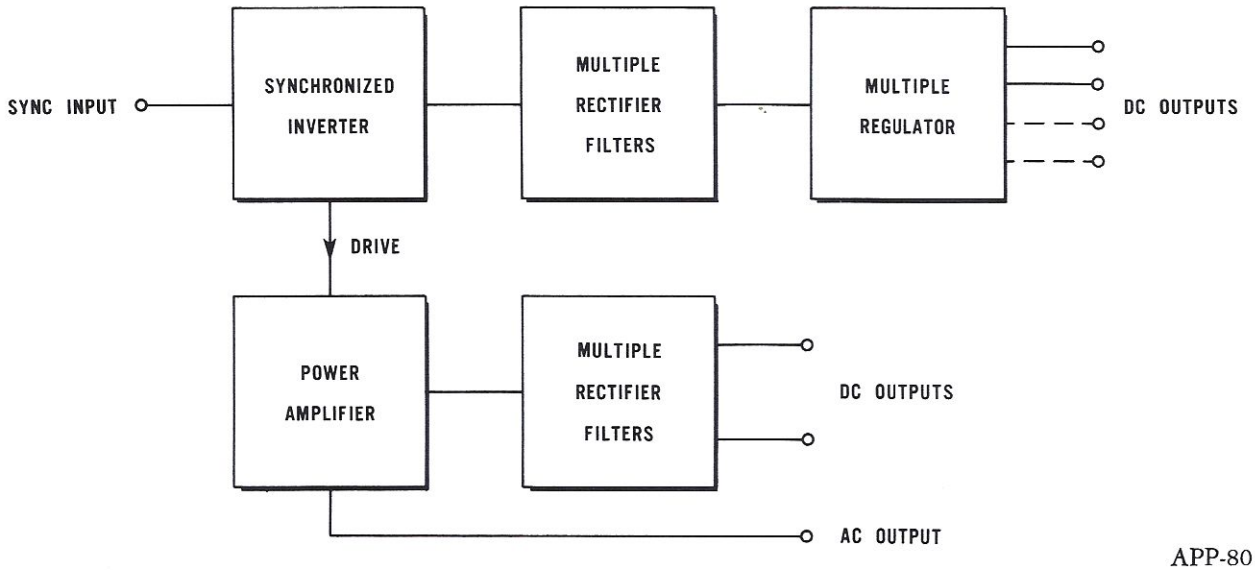

FIG. 1. Block diagram of power supply.

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in a synchronized power-oscillator circuit and two in a power amplifier driven by the oscillator. 2N2892, 2N1893, and 2N915 transistors are used in the series regulator circuits on the output side of the converter.

II. DESCRIPTION OF POWER-SUPPLY CIRCUIT

Figure 1 shows a block diagram of the power supply. Approximately half the total power required by the radar is supplied by the output from the synchronized oscillator, and the other half by the output from the power amplifier. The a-c output supplying the motor load was taken from the power amplifier rather than the synchronized oscillator, so that the motor loads would not impair synchronization.

The synchronized oscillator, designed to run more slowly than the synchronized frequency of 465 cps, will lock in with the synchronization pulses. The sync input frequency is 930 cps, the same as the prf of the radar. This frequency, being twice the oscillator operating frequency, causes switching on each half cycle and this results in perfect symmetry of the output voltage waveshape and prevents any imbalance in output transformer primary d-c current which could lead to undesirable core saturation.

The output from the oscillator feeds five sets of rectifier-filters and, with their accompanying regulator circuits, provides the following d-c outputs:

1. Bias bus voltage, -6 V at 20 mA.
2. Collector supply voltage, +12 V at 0.5 A.
3. Klystron shell voltage, +300 V at 30 mA.
4. Video amplifier collector voltage, +70 V at 5 mA.
5. Modulator voltage, +500 V at 35 mA.
6. TR tube keep-alive voltage, -400 V at 100 μA.
7. Klystron repeller voltage, -150 V at 1.0 μA.

Two additional low-voltage a-c outputs are taken from the synchronized oscillator to drive the base circuits of the power amplifier. The amplifier transistors, like the oscillator transistors, operate in the switching mode, and therefore dissipate minimum power. The power amplifier, which has one output feeding two sets of rectifier-filters and one unrectified output, provides the following supply voltages to the radar:

1. Heater voltage for magnetron and time delay relay, 8 V d-c at 0.6 A.
2. Heater voltage for the klystron, indicator CRT and graticule lamps, 6.3 V d-c at 2 A.
3. Voltage for antenna and indicator drive motor, 26 V a-c. (The square-wave voltage powers the motors without being filtered into a sine wave.)

The high-voltage supply for the PPI cathode-ray tube is located in the indicator unit separate from the remainder of the power supply in the transmitter unit. This voltage supply consists of a transformer and a rectifier-filter operating from the same a-c voltage that drives the motors.

For operation from a 28 V d-c source Fairchild 2N2892 power transistors are used in the synchronized oscillator and power amplifier. If operation from a 12 V d-c source is desired, higher-current devices, either silicon or germanium, are recommended as power switches and the oscillator transformer in the synchronized oscillator must be of special design. However, the power transformers in the synchronized oscillator and in the power amplifier need not be changed for lower-voltage operations, since taps are provided on their primary windings for connections to the lower-voltage power transistors. Thus all circuitry on the secondary side of the transformer remains unchanged for the different input voltages.

For operation from 115 V a-c a suitable transformer, rectifier and filter combination should be

provided to supply an output of 28 V d-c to the basic oscillator-amplifier circuit. If line-voltage changes of greater than ±10% are anticipated, a regulator should be incorporated at the output of the filter to hold the 28 V d-c supply to within ±10%. Such a regulator is advantageous as it also suppresses, to a large degree, the power-supply ripple and thus simplifies the power-supply filter design.

The schematic diagram of the power supply is shown in Fig. 2.* The synchronized oscillator comprising transistors Q_1 and Q_2 utilizes a low-power oscillator transformer T_1, which determines the non-synchronized or free-running operating frequency. Power is taken from the oscillator through the power transformer T_2. Although transformers T_1 and T_2 are connected in parallel, their functions are different.

When power is first applied to the oscillator circuit, the voltage drop across diode D_2 due to the current flow through resistors R_1 and R_2 forward-biases the oscillator transistors Q_1 and Q_2 enough to cause collector current flow. The positive feedback from the emitter-to-base circuit through transformer T_1 causes the circuit to oscillate. The base current is limited by resistor R_1, and a return current path exists through diode D_1. Synchronization is effected by negative-going sync pulses introduced into the bases of transistors Q_1 and Q_2 through diodes D_3 and D_4. Transistors Q_{10} and Q_{11} provide trigger-pulse inversion and amplification, respectively, for effecting the synchronization action, while L_1 provides the correct magnetic condition to permit synchronous-mode operation of transformer T_1.

If the power leads from the power source are long enough that appreciable inductance is introduced between the oscillator and the power source, fast current pulses in the oscillator produced by the switching action would develop excessively large transient voltages at the input power terminals. As this voltage could affect the transistors adversely, capacitor C_1 is placed across the terminals to integrate the current pulse and thus keep the terminal voltage essentially constant.

*See fold-out back page for Fig. 2.

The low voltage bias and collector supply power for the transistor circuitry in the radar are taken from a single output winding on transformer T_2. The full-wave bias rectifiers consist of diodes D_5 and D_6, and capacitor C_3 provides filtering. A zener diode D_9 acts as a low-power shunt regulator to hold the output voltage constant at approximately -6 volts. The collector supply has full-wave rectification provided by diodes D_7 and D_8 and filtering by capacitor C_4. To offset the large current drain on the +12 V collector supply, a series regulator circuit with feedback is required rather than a simple shunt zener regulator as used in the bias supply. The series regulator is a Fairchild 2N2892 power transistor driven by an emitter-follower utilizing a Fairchild 2N1613. The error amplifier within the feedback loop uses a Fairchild 2N915 and the voltage reference is provided by zener diode D_{10}. Both the collector current for transistor Q_5 and the base current for transistor Q_6 are obtained from the current source comprising resistor R_5 and the +300 V supply. The output voltage is adjustable to the proper value of 12 V by potentiometer R_6.

The +300 V supply power is taken from a high-voltage winding on transformer T_2 and rectified by diodes D_{11} and D_{12} with filtering accomplished by capacitor C_6. The series regulator comprising transistor Q_9 is driven by transistor Q_8 in a non-feedback circuit. The output voltage is equal to the voltage applied to the base of transistor Q_8 minus the sum of the base-to-emitter voltages of Q_8 and Q_9. The base voltage for Q_8 is adjustable by potentiometer R_9, and the voltage across R_9 is maintained constant by the shunt-regulator zener diode D_{13}. The current source for the zener diode is provided by resistor R_{10} operating from the +500 V supply.

The +70 V regulator, which supplies low output current, contains a shunt zener diode, D_{14}. The source for the 70 V regulator is the output of the 300 V regulator.

The power required for the transmitter is taken from a second high-voltage winding on transformer T_2. A full-wave rectifier is used consisting of two sets of two series-connected diodes. This arrangement provides the high PIV rating required for the rectifiers in this circuit. Capacitor C_{10} is the filter capacitor. No regulation is provided in the circuit. An additional winding added to the previous high-voltage winding provides the voltage for the TR keep-alive supply. Half-wave rectification by diodes D_{19} and D_{20} produces a -700 V output filtered by capacitor C_{11}. Resistor R_{15} added to the -700 V supply provides a current source for the TR tube, and as a result the terminal voltage at the supply due to the drop across R_{15} under the TR tube load is -400 V. The klystron repeller voltage is obtained by operation of a 150 V shunt regulator comprising zener diode D_{21} from the -700 V output. Adjustable output is provided by potentiometer R_{13}.

The base-drive power for the power-amplifier transistors Q_3 and Q_4 is obtained from two low-voltage windings on transformer T_2. Base-current limiting for transistors Q_3 and Q_4 is provided by resistors R_{17} and R_{18}, respectively. Thus the voltage across the primary winding of transformer T_3 has the same amplitude and waveshape as that across the primary of transformer T_2.

The +8 V magnetron heater power is obtained from a low-voltage secondary winding of transformer T_3. Although the magnetron heater requires nominally 6.3 V, the voltage drop in the modulator pulse transformer, through which the heater current flows, necessitates an 8 V supply. Full-wave rectification is accomplished by diodes D_{22} and D_{25} and filtering by capacitor C_{13}. A similar rectifier-filter, comprising diodes D_{23} and D_{24} and capacitor C_{14}, operating from taps on the low-voltage secondary winding, results in a +6.3 V output for tube heaters. An additional low-voltage winding on transformer T_3 furnishes a-c power to the motors for antenna and indicator rotation. A small amount of power is also taken from the winding for powering the high-voltage supply for the indicator CRT.

III. DESIGN OF SYNCHRONIZED OSCILLATOR

The synchronized oscillator must have a non-synchronized free-running frequency of less than the sync pulse rate over the full range of input line voltage variation. This variation will nominally be ±10%; that is, with 28 V as the normal line voltage, the input voltage may vary from approximately 25 to 31 V d-c. The frequency of the magnetically coupled multivibrator is directly proportional to voltage so that if a frequency of 430 cps is chosen to correspond to the upper limit of 31 V d-c input, the frequency for the lower limit of 25 V d-c input will be 347 cps. The power transformers in both the synchronized oscillator and the power amplifier must operate as linear, nonsaturating magnetic elements in the above frequency range. The volt-second area absorbed by the transformer core-coil combination is the same at 25 V d-c with 347 cps as it is at 31 V d-c with 430 cps. Therefore the transformer design can be carried out using the numerical values at either limit. The same transformer design can also be used for the synchronized condition, where the frequency is increased to 465 cps, since the operating flux density then decreases slightly from its nonsynchronized value.

The oscillator transformer operated in a switching circuit determines the frequency of operation according to the following formula:[1]

$$f = \frac{E \times 10^8}{4 N_p \phi_s}$$

The number of primary turns on each side of the primary center tap is determined from the above formula as follows:

$$N_p = \frac{E \times 10^8}{4f\phi_s} = \frac{E \times 10^8}{4fB_sA_c}$$

A suitable transformer core material is 6 mil Orthonol[2] wound into a toroid having an O.D. of 1.50", an I.D. of 1", and a height of 0.25" with a resulting cross-sectional area A_c or 0.363 cm^2. The flux density at which saturation becomes pronounced is approximately 13 kilogauss.

$$N_p = \frac{31 \times 10^8}{4\,(430)\,(13 \times 10^3)\,(0.363)}$$

The above calculation assumes that the voltage drop across the switching transistors in the on condition is zero and therefore the induced voltage across the primary is equal to the supply. Since the switches also carry load current, however, there is a voltage drop across them on the order of 1 V, which reduces the induced voltage, and hence the frequency, by approximately 3%. This reduction produces no objectionable effect in either the free-running or the synchronized mode.

The feedback voltage is set to a value of 4 V, which allows for the forward base-to-emitter drop of the transistors and an additional voltage which, in combination with the base current-limiting resistor, provides a current-drive source for the transistor bases. The feedback winding is determined as follows:

$$N_{FB} = \frac{E_{FB}}{E_{pri}} \times N_p = \frac{4}{31} \times 382 = 49 \text{ turns}$$

The oscillator transformer and the associated inductor specifications are shown in Fig. 3.

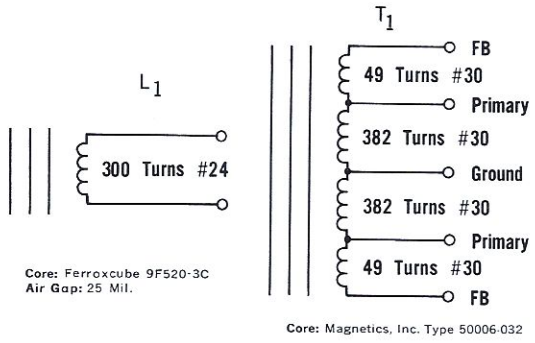


FIG. 3. Winding data on oscillator transformer.

The base current-limiting resistor R_1 is designed to allow a maximum collector current flow I_C greater than 1.5 A, which is the sum of the magnetizing current plus the load current. If I_C is set at 3 A, the value of R_1 may be calculated as follows:

$$R_1 = \frac{E_{FB} - V_{BE}}{I_{C(max)}/\beta}$$

$$E_{FB} = 4\,V,\ V_{BE} = 1\,V,\ I_{C(max)} = 3\,A,\ \beta = 25$$

$$R_1 = \frac{4 - 1}{3/25} = 25\,\Omega$$

The maximum current through the winding is the sum of base drive current and the magnetizing current and totals approximately 150 mA. Since the operating duty cycle of each primary and feedback winding is 50%, the average current is 75 mA. For a transformer of this small size, a winding wire with a current-carrying capacity of 400 circular mils per ampere is satisfactory. Therefore the wire area need be no larger than 0.075 x 400, or 30 circular mils, which corresponds to AWG No. 35. However, since the core could accommodate a larger copper volume, No. 30 wire was used in the transformer because it was easier to wind.

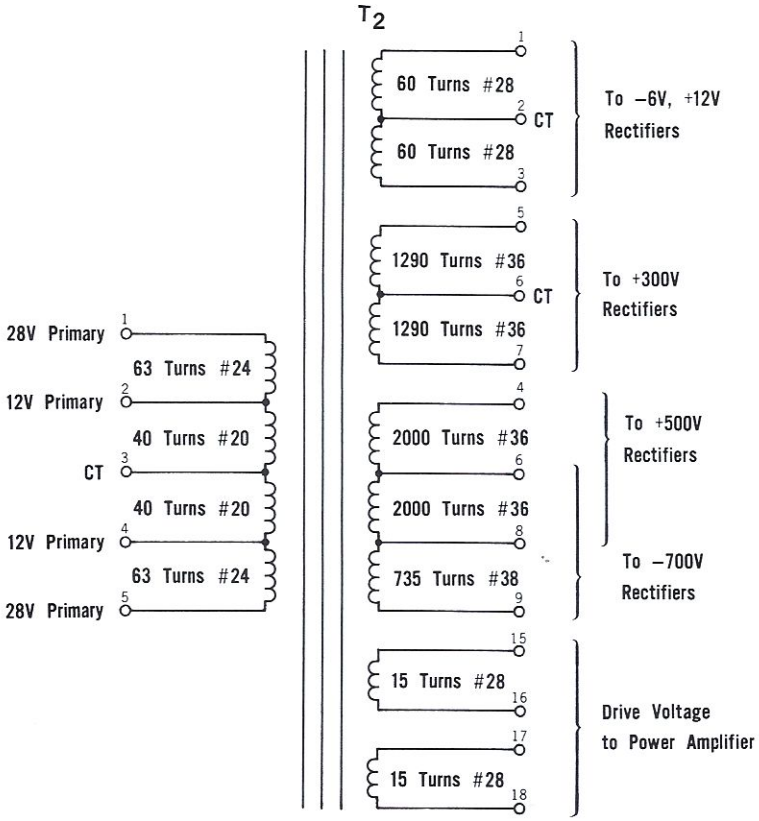
The converter power transformer is wound on a 6-mil Orthonol core and operated at a maximum flux density $B_{(max)}$ of 12 kilogauss. The primary turns can be determined by the same formula as the oscillator transformer.

$$N_p = \frac{E \times 10^8}{4f\phi_{(max)}} = \frac{E \times 10^8}{4fA_cB_{(max)}}$$

The particular core has an O. D. of 2", an I. D. of 1.5", and a height of 1", with a cross-sectional area of 1.452 cm^2.

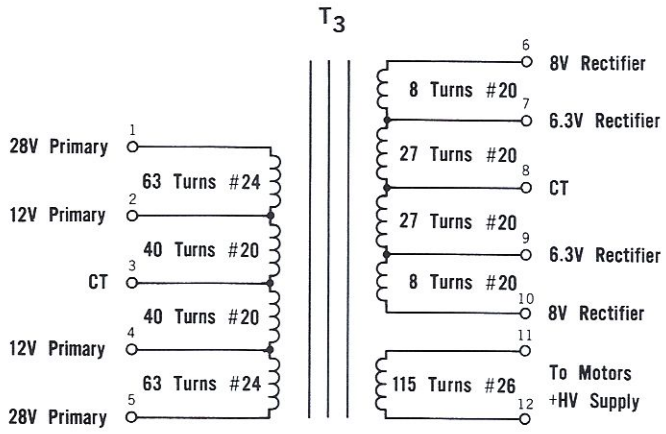
$$N_p = \frac{31 \times 10^8}{4\,(430)\,(1.452)\,(12 \times 10^3)} = 103 \text{ turns}$$

The same transformer can also be driven from a 12 V square wave by tapping the winding at $\frac{12}{31}$ (103) or 40 turns. As shown in Fig. 4 the wire size for each winding of the converter power transformer is based on 500 circular mils per ampere. The secondary turns are based on a nominal turns-per-volt ratio of 103/28, or 3.68. An additional 3% of turns added to each winding allows for transformer regulation. Care must be exercised in both the placement of the turns and insulation on the high-voltage secondaries to prevent breakdown.



Core: Magnetics, Inc. Type 50006-040
Winding Order:
Primary windings 1-2-3-4-5 are placed next to core; windings 15-16,17-18 are placed over primaries; windings 1-2-3 are then added; high voltage insulation placed over all the preceding low voltage windings; remaining high voltage secondaries are added, first winding 5-6-7, then winding 4-6-8-9.

FIG. 4. Winding data on converter power transformer.



Core: Magnetics, Inc. Type 50006-040
Winding Order:
Primary windings 1-2-3-4-5 are placed next to core; windings 6-7-8-9-10 are placed over primaries; winding 11-12 is then wound.

FIG. 5. Winding data on power amplifier transformer.

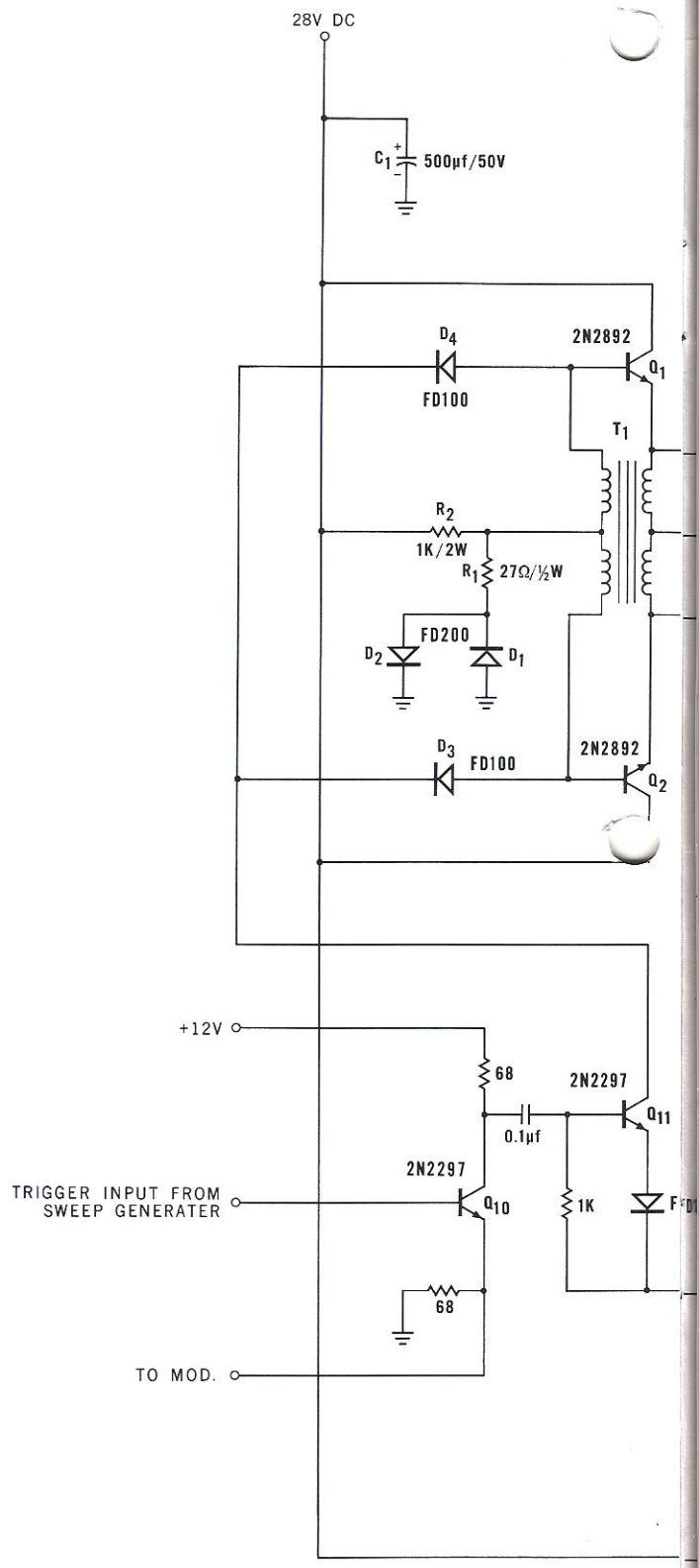
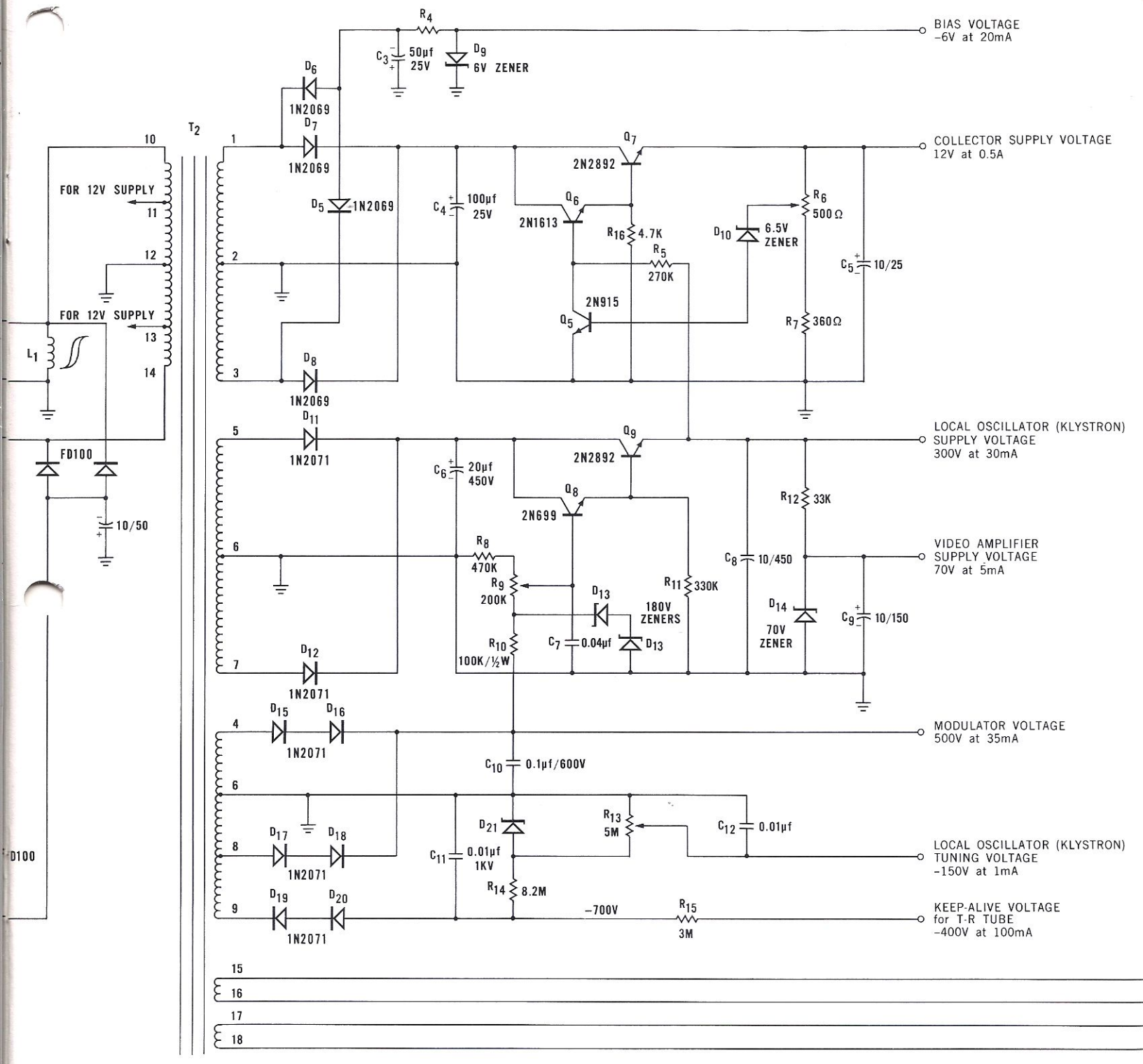


FIG. 2. SCHEMATIC DIAGRAM OF POWER SUPPLY



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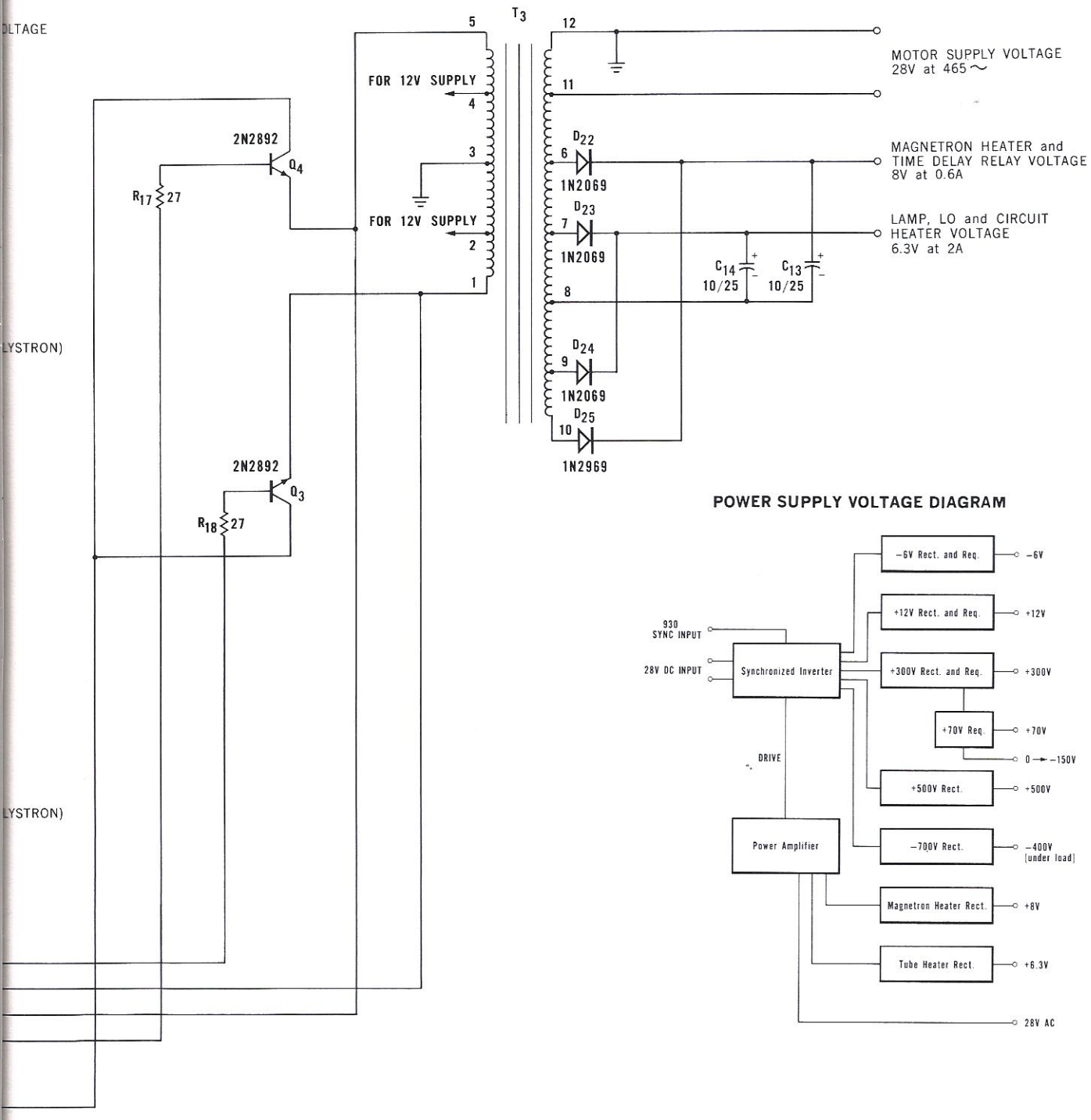

POWER SUPPLY VOLTAGE DIAGRAM

The power transformer in the driver power amplifier, which has approximately the same volt-ampere rating as the converter power transformer, is wound on the same core and has an identical primary winding. The secondary turns are based on the same turns-per-volt and regulation requirement as the converter power transformer. Data on the power transformer are given in Fig. 5. The values of the base current-limiting resistors are also identical to those used in the oscillator converter circuit.

IV. DESIGN OF THE 12 V D-C REGULATOR

By means of a series regulator the collector supply voltage for the radar circuitry is maintained within ±2% for ±10% changes in input supply voltage and only negligible changes in load current. A transistor provides a nearly ideal series-regulator element since its high collector resistance tends to absorb large collector voltage changes with very small emitter current change. Thus an adequate regulation capability may be obtained with a single series transistor whose base is maintained by a constant voltage source. In order to provide control capability of the output voltage, however, a feedback regulator employing a single error-voltage amplifier proves to be the most practical arrangement.

As shown in Fig. 6, the series-regulator transistor Q_7 is driven by another similarly connected transistor Q_6, and as a result the drive-current requirement of the error amplifier is reduced. Since the base of transistor Q_7 is returned to ground through resistor R_{16}, I_{CBO} effects are prevented when the power supply is tested at high temperature with no load. A satisfactory value for R_{16} is 4.7 KΩ, which allows for maximum I_{CBO} in addition to an extra margin of approximately 2.7 mA. With a low-limit β of 30 for transistor Q_7, the required base current is 0.5/30, or approximately 17 mA. The resulting emitter current of transistor Q_6 is 17 + 2.7, or 19.7 mA, and with a low limit β of 40 the necessary base current is approximately 0.5 mA.

The current through the control amplifier Q_5 is selected at 0.5 mA so that resistor R_5, which supplies this current plus the base current of Q_6, can be calculated as

$$\frac{300 - 12}{1 \times 10^{-3}} = 288 \text{ K}\Omega$$

The base current of transistor Q_5 is

$$\frac{I_C}{\beta_o} \approx \frac{0.5}{30} \text{ mA} \approx 0.17 \text{ mA}$$

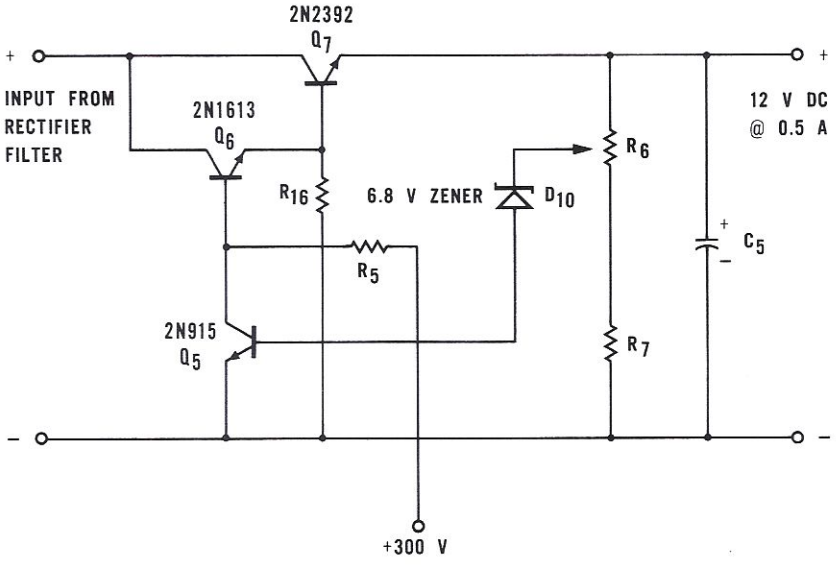

FIG. 6. Schematic diagram of 12 V d-c regulator.

The value of the feedback network consisting of adjustment potentiometer R_6 and resistor R_7 is selected so that the current through it is at least ten times as large as the base current of Q_5, and hence the regulation of the feedback network itself is held at a small value. The zener diode is a base-emitter diode of a 2N915 transistor and has a nominal zener voltage of 6.8 V. Capacitor C_5 is added to the circuit to insure stability against high-frequency oscillations within the regulator.

The voltage from collector to emitter of transistor Q_7 goes from 2 V to 5 V as the input supply voltage varies $\pm 10\%$. The worst-case dissipation in Q_7 occurs at high input voltage and is 5 V x 0.5 A, or 2.5 W. Adequate heat sinking is obtained by mounting Q_7 to the chassis with insulating hardware. Similarly Q_6 has a worst-case dissipation of 5 V x 19.7 mA, or 98.5 mW and requires no heat sink.

ACKNOWLEDGEMENT

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2. Core manufactured by Magnetics, Inc., Butler, Pennsylvania, Type 50006 - 032.